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TITLE: A Novel Approach for Bandwidth Enhancement of Slot Antennas

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TITLE: Proceedings of the 2003 Antenna Applications Symposium [27th]
Held in Monticello, Illinois on 17-19 September 2003. Volume 1

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A NOVEL APPROACH FOR BANDWIDTH ENHANCEMENT OF SLOT ANTENNAS

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Abstract: A novel approach is presented to improve the bandwidth of slot antennas. The technique is based on manipulating the field distribution along an ordinary resonant slot structure using the feed line and creating a dual resonance behavior. Hence without changing the length of the antenna its bandwidth is increased by more than 200% relative to a narrow slot. The field distribution along the slot at a frequency slightly above its natural resonance is manipulated by a narrow microstrip line feeding the slot near one of the two edges. By proper choice of the slot width, feed location, and microstrip feed line a fictitious second resonance can be created by establishing a null in the electric field distribution along the slot near the feed line. This null is resulted from the superposition of the microstrip near field and the slot field excited by the displacement current. A prototype is designed and tested at the center frequency 3.4 GHz. A large bandwidth of 37% is achieved without any constraints on impedance matching or complexity in the antenna structure. Also bandwidth enhancement of a miniaturized slot antenna using parasitic coupling is presented. The antenna occupies a small area of $0.15\lambda_0 \times 0.15\lambda_0$ and can have up to 3% bandwidth.

1. INTRODUCTION

High bandwidth, small size, simplicity, and compatibility to the rest of the RF front-end are desirable factors of an antenna. Enormous effort has been invested on designing frequency independent or very wide band antennas. One of the major drawbacks of such antennas is their relatively large size which can potentially eliminate their use for mobile wireless applications. Therefore, it is desirable to develop other techniques to increase the bandwidth of otherwise narrowband antennas without significantly increasing their sizes. Here we describe two different techniques for enhancing the bandwidth of slot antennas. Microstrip-fed wide slot antennas have been theoretically studied in [1] and experimental investigation on very wide slot antennas that result in wide band antennas has also been performed by various authors [2], [3] but not much effort

has been put into the investigation of the radiation characteristics of the moderately wide slot antennas. Unlike very wide slots, these antennas do not produce high levels of cross polarization and their feeding mechanism is much simpler than the fork shaped [2], [3] or T-shaped microstrip feed [4] normally used in very wide slot antennas. In the following two sections a microstrip-fed wide slot antenna is studied and it is shown that its bandwidth can be more than doubled by creating a fictitious short circuit along the slot or it can be operated in a dual band mode with considerable bandwidth at both frequency bands.

The fundamental limitations on electrically small antennas have been extensively studied by various authors [6]-[8]. Early studies have shown that for a single resonant antenna as the size is decreased its bandwidth (BW), if it can be matched, and efficiency decrease too [6]. This is a fundamental limitation which in general holds true independent of antenna architecture. Recently, there have been a number of studies on different approaches for antenna miniaturization while maintaining a relatively high bandwidth and efficiency. A novel miniaturized slot antenna was recently presented [9]. This element shows good radiation characteristics but has a rather small bandwidth (less than 1%). In the second part of this paper Bandwidth enhancement of this class of miniaturized antennas using a dual resonant topology is examined. Basically the miniaturized slot antenna configuration similar to the one used in [9] is considered to form a dual antenna structure. In this new configuration one of the elements is fed by a microstrip transmission line and the other one is fed parasitically by the first antenna. It is shown that the bandwidth of this new double-element configuration is twice the bandwidth of a single antenna that occupies the same space. In other words by keeping the same space as the single antenna the bandwidth is increased by a factor of two.

In what follows first the design of a microstrip-fed broadband slot antenna is presented in section 2 then dual band characteristics of such antennas is examined in section 3 and in section 4 bandwidth enhancement of miniaturized antennas using parasitic coupling will be studied.

2. WIDEBAND MICROSTRIP-FED SLOT ANTENNA

A resonant narrow slot antenna is equivalent to a magnetic dipole at its first resonant frequency. Its length is $\lambda_g/2$ where λ_g is the guided wavelength in the slot. If the slot antenna is fed near an edge by a very narrow microstrip line and the slot width is properly chosen, at a frequency above the first slot resonance, a fictitious short circuit near the microstrip feed may be created. Basically the tangential electric field created by the narrow microstrip line at a particular distance cancels out the electric field of the slot excited by the return current on

the ground plane. A full wave simulation shows the field distribution for this situation in Figure 1(b) whereas Figure 1(a) shows the electric field distribution in the slot at its normal first resonance. The field distribution is shown to have a null along the slot between the two ends at a frequency slightly above the slot resonance.

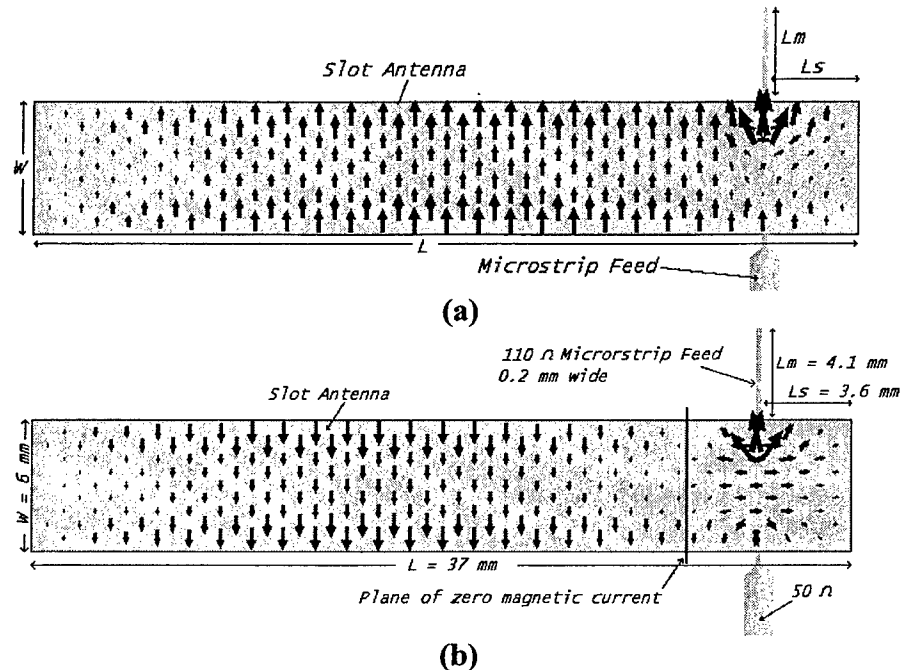


Figure 1. Electric field distribution along a wide slot antenna at the first resonance frequency (a) and the fictitious second resonance (b). (The figures are obtained from full wave simulations)

The width of the microstrip feed, width of the slot, and the distance between the feed and the edge of the slot antenna, L_s , are parameters that affect the existence and location of this fictitious short circuit. As L_s increases, the second resonance frequency also increases therefore by choosing L_s the second resonance frequency can be chosen such that the total antenna bandwidth is increased or a dual band operation is achieved. Matching is achieved by tuning length of the open circuited microstrip line, L_m . Since L_s is being set by the second resonance frequency there is not much flexibility in obtaining a good match by changing the length of L_s .

As can be seen from Figure 1(b), the electric field distribution at this second resonance frequency is similar to that of the first one therefore, it is expected that the radiation patterns of the antenna at the two frequencies be similar. The wideband slot antenna has a length of $L=37\text{mm}$ and width of $W=6\text{mm}$ and was simulated using IE3D [10] and fabricated on a $500\mu\text{m}$ thick, RO4350B substrate with dielectric constant of 3.4, $\tan \delta=0.003$ and a ground plane area of $15\text{ cm} \times 11\text{ cm}$. Figure 2 shows simulated and measured return losses of this antenna. Two

different antennas were fabricated and the measured return losses of both of them are shown in Figure 2. The only difference between the two antennas is in the location of the feed. In Antenna #2 L_S is longer than Antenna #1 by 1mm therefore the frequency of second null in S_{11} is higher and the overall bandwidth is larger. The -10 dB bandwidth of Antenna #1 is 1022 MHz (30.3%) and that of Antenna #2 is 1220 MHz (37%). The slight discrepancy between the simulation and measurement results can be attributed to the finiteness of the ground plane which causes a shift in the resonance frequency and the fact that the response of the system is very sensitive to the exact location of the microstrip feed which is subject to alignment errors in the implementation process.

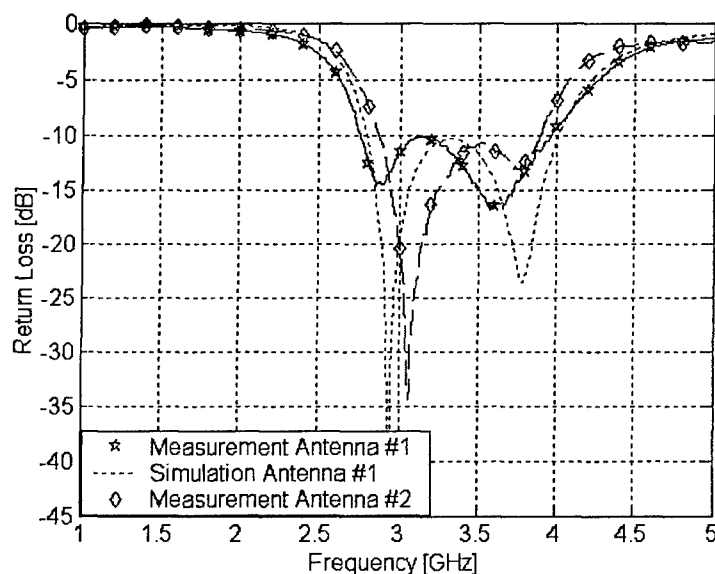


Figure 2. Return losses of the wideband slot antennas.

Radiation patterns of the antenna were measured in the anechoic chamber of the University of Michigan. Figure 3 shows the E- and H-plane co- and cross-polarized patterns at 3.077 GHz and 3.790 GHz. It is shown that the patterns are almost dual of those of an electric dipole of the same length. The deeps in the E-plane at $\pm 90^\circ$ are caused by the out of phase radiation from the edge of the substrate. It is also seen that the cross polarization level is negligible.

At the second resonance frequency, in addition to the radiation from the edges of the ground plane the oppositely directed magnetic current in one section of the slot can also contribute to the formation of the deep in E-Plane pattern at $+90^\circ$ therefore there will be a deeper null in E-Plane at an azimuth angle of $+90^\circ$.

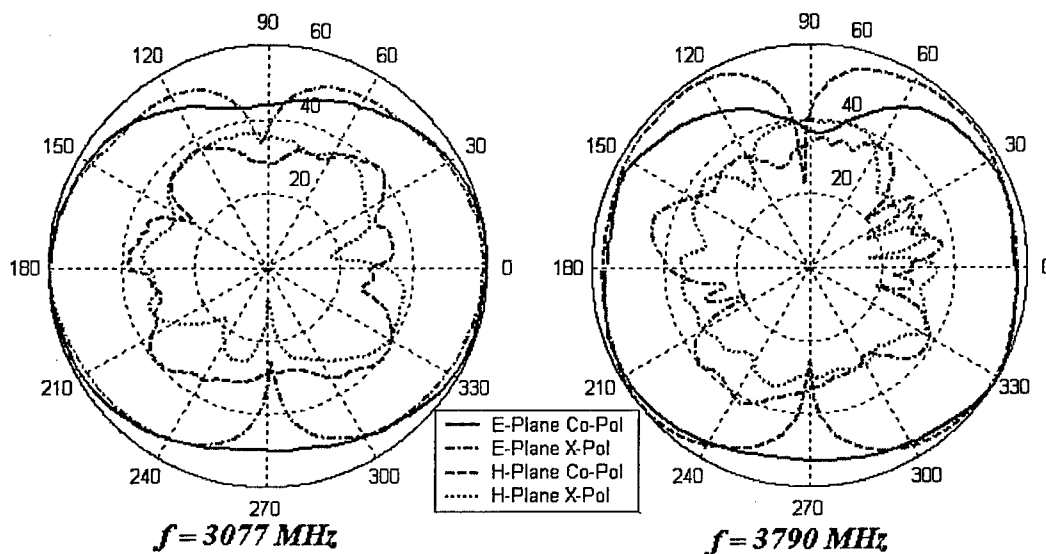


Figure 3. Radiation patterns of the wideband antenna measured at the two resonance frequencies.

The antenna gains were measured at the two resonance frequencies using a reference double ridge horn antenna and are reported in Table I. It is important to note that at f_u the maximum radiation does not take place at bore-sight therefore the gain at bore-sight is lower than the maximum antenna gain.

Type	L, W [mm]	Ls, Lm [mm]	Bandwidth	Gain* @ f_l, f_u
Antenna 1	37, 6	3.6, 4.1	30.3%	2.5 dB, 0.1 dB
Antenna 2	37, 6	4.6, 3.9	37.0%	2.5 dB, 0.0 dB

TABLE I. PHYSICAL AND RADIATION PARAMETERS OF THE BROADBAND SLOT ANTENNA

* Gain is measured at bore-sight.

3. DUAL BAND MICROSTRIP FED SLOT ANTENNA

The wide slot antenna has also the capability of operating in a dual band mode. If the distance L_s is increased the equivalent length of the second resonance decreases therefore the second resonance frequency increases while the location of the first resonance does not change considerably, because it is being set by the overall length of the antenna, therefore the antenna becomes a dual band one. The separation between the two resonance frequencies is a function of the resonant length of the second frequency and as a result of it a function of the distance L_s . Figure 4 shows the simulation results of thick slot antenna with $L=31$ mm and $W=6$ mm on the RO4350B substrate for different values of L_s .

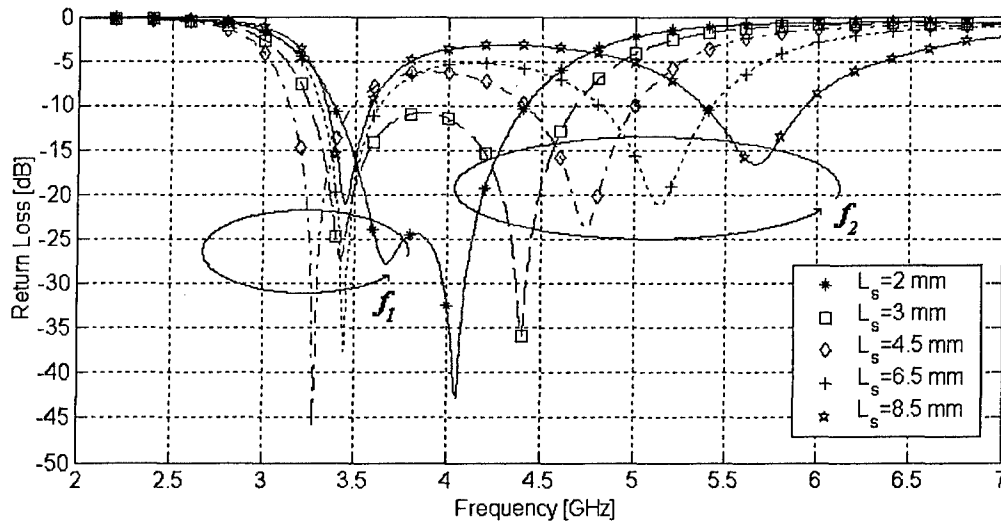


Figure 4. Simulation results of the return loss of the wide slot antenna for different L_s values.

As is observed from this figure, by increasing L_s from 2mm to 8.5 mm, f_1 varies between 3.7 GHz and 3.3 GHz and f_u increases from 4GHz to 5.2 GHz. It can be seen that f_u/f_1 ratios of 1.6 or more can easily be obtained from this antennas.

The magnetic current distribution at these two different frequencies are similar to each other and result in a similar radiation patterns at the two bands which is expected in a dual-band operation. Figure 5 shows the magnitude of the electric field distribution along the slot for the two resonance frequencies for the case of $L_s=8.5$ mm. A dual band antenna with $L=31$ mm, $W=6$ mm and $L_s=5$ mm was fabricated on the same RO4350B substrate. The simulated and measured return losses of this antenna are shown in Figure 6. As can be observed, an excellent agreement is observed between the simulation and measurement results. The antenna has f_1 and f_u of 3.3 GHz and 4.85 GHz respectively. The bandwidths of the antenna in the first and second bands are 10.6% and 12.1% respectively. It should also be noticed that the antenna has considerable bandwidth in both frequency bands which makes it an excellent choice for wireless communication applications.

The radiation patterns of the antenna were measured in the anechoic chamber of the University of Michigan and the E- and H-Plane, co- and cross-polarized radiation patterns are presented at Figure 7. It is observed that the radiation patterns at both frequencies are similar to each other and the cross polarization level is negligible. As can be seen from Figure 7(b), the maximum radiation in the

H-Plane does not occur at bore-sight but rather at around $\phi=\pm 20^\circ$ for both frequency bands.

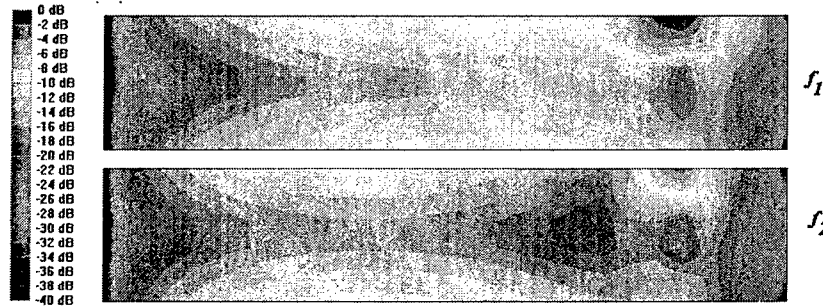


Figure 5. Magnitude of the electric field distribution of the dual band slot antenna at the two resonance frequencies.

L, W [mm]	L _s , L _m [mm]	f_l , f_u [GHz]	BW @ f_l , f_u	Gain* @ f_l , f_u
31, 6	5, 2.3	3.3, 4.85	10.6%, 12.1%	0.5 dBi, 0.3 dBi

TABLE II. A SUMMARY OF THE PHYSICAL AND RADIATION PARAMETERS OF THE DUAL BAND SLOT ANTENNA

* Gain is measured at bore-sight.

The antenna gain at bore-sight is measured using a standard double ridged horn antenna and is reported in Table II. The maximum gain of the antenna in the H-Plane occurs around $\phi=\pm 20^\circ$ and is measured to be 2.5 dB and 3.0 dB for f_l and f_u respectively. The efficiency of the antenna can be calculated using the measured values for gain and calculated directivity values. Based on this method efficiency values of 91% and 89% are obtained for the two frequency bands.

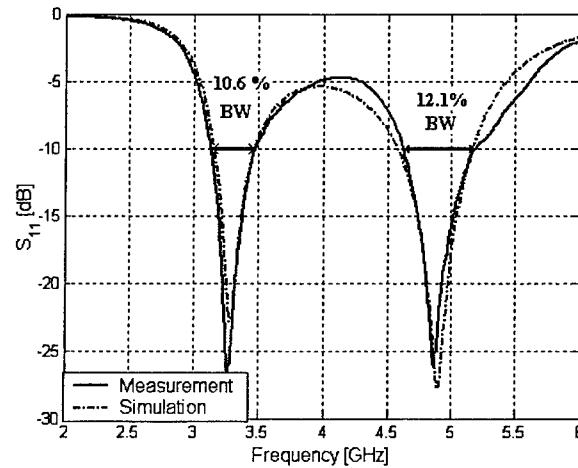


Figure 6. Simulated and measured return losses of a dual band antenna with $L=31\text{mm}$, $W=6\text{mm}$ and $L_s= 5\text{mm}$.

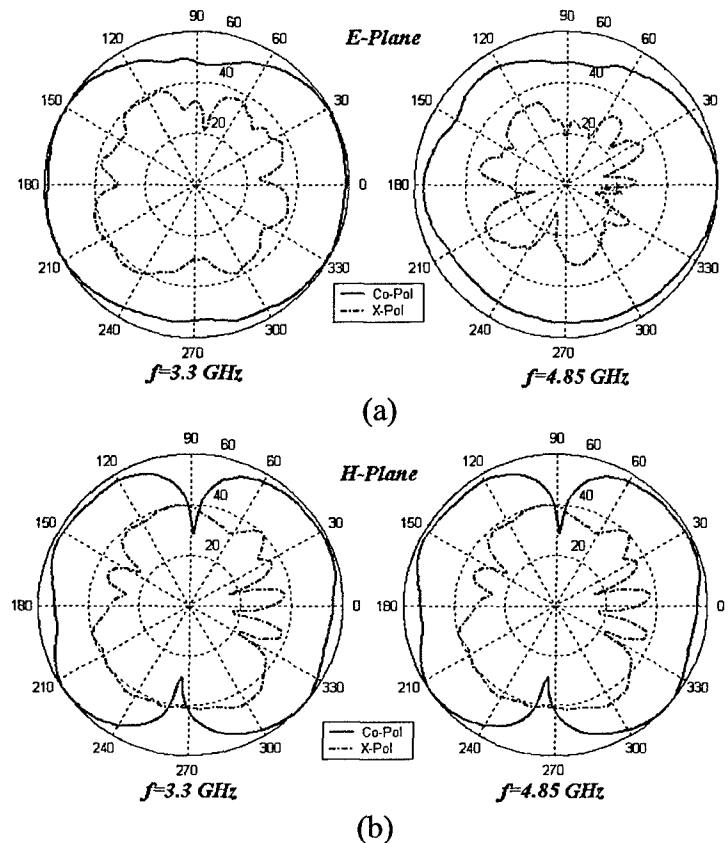


Figure 7. Measured radiation patterns of the dual band slot antenna at $f_1=3.3$ GHz and $f_2=4.85$ GHz. (a) E-Plane (b) H-Plane

4. PARASITICALLY COUPLED MINIATURIZED SLOT ANTENNA

In this section we investigate the design of a double-element miniaturized antenna for bandwidth enhancement. This miniaturized antenna occupies an area of $0.15\lambda_0 \times 0.13\lambda_0$ but shows a rather small bandwidth (less than 1%) [9]. A close examination of the antenna topology reveals that the slot structure covers only half of this rectangular area, and therefore another antenna with the same geometry can be placed in the remaining half without significantly increasing the size (Figure 8).

Placing two antennas in such proximity to each other creates significant coupling which can be taken advantage of to increase the total bandwidth of the antenna. In this design the two antennas are tuned to resonate at $f_0=850$ MHz by adjusting the overall slot length to an effective length of $\lambda_g/2$, where λ_g is the guided wavelength in the slot.

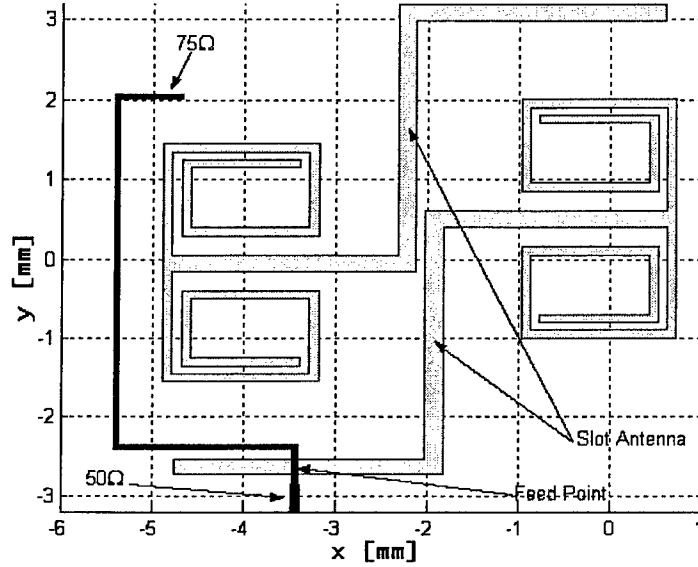


Figure 8. Geometry of the double-element miniaturized slot antenna.

This way, the structure acts like a coupled resonator the bandwidth of which is a function of coupling coefficient between the two resonators. By considering this antenna as a two port network where the second port is in free space, its behavior can be explained using the coupled resonator filter theory. The resonant frequency of the antenna (f_0) determines the center frequency of the operation. The frequency domain reflection coefficient (S_{11}) of this second order system has two distinct zeros with a separation proportional to the coupling coefficient. Bandwidth maximization is accomplished by choosing a coupling coefficient such that S_{11} remains below -10 dB over the entire frequency band as the two zeros are separated. The coupling coefficient which can be controlled by changing the length of the overlap section and the separation between the two antennas (d and s in Figure 10) and is defined as:

$$k_t = \frac{f_u^2 - f_l^2}{f_u^2 + f_l^2} \quad (1)$$

Where, f_u and f_l are the frequencies of the S_{11} nulls and $f_u > f_l$. The coupling is a mixed electric and magnetic coupling but the electric coupling is stronger in the middle section of both antennas (Figure 10) where the separation, s , is small and the E-Field is large therefore it is the dominant coupling mechanism which results in in-phase magnetic currents in both slot antennas. Figure 9 shows that the two zeros occur at 848MHz and 860 MHz which corresponds to $k_t=0.014$.

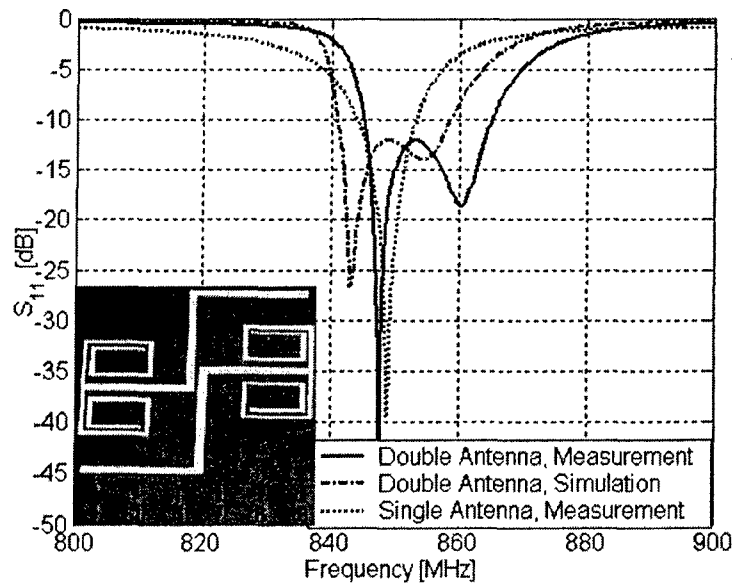


Figure 9. Return losses of the double-element miniaturized antenna and a single element antenna that constitutes it

The input resistance of a microstrip-fed slot antenna depends on the location of the microstrip feed relative to the slot and varies from zero in the short circuited edge to a high impedance at the center. Therefore an off center feed can be used to match the antenna. Here, the feed is a narrow 75Ω microstrip open circuited transmission line connected to a main 50Ω microstrip line. The length of the open circuited microstrip line can be chosen to compensate for the reactive part of the input impedance. Here the input impedance is not purely real therefore the 75Ω line is extended by $0.33\lambda_m$ (λ_m is the wavelength in microstrip line) after the strip-slot crossing to compensate for the reactive part of the input impedance.

The double-element slot antenna was simulated using *IE3D* [10] and fabricated on the $500\mu\text{m}$ thick Rogers RO4350B substrate. Figure 9 shows the measured and simulated return losses of the double-element antenna and the measured return loss of a single element that constitutes it. The minute discrepancy between the simulated and measured results can be attributed to the finite size of the ground plane. The antenna shows a -10dB bandwidth of 21 MHz or 2.4% which is more than twice (2.7 times) the bandwidth of a single element antenna (8 MHz or 0.9%). This bandwidth can also be further increased (up to 3%) by increasing the coupling coefficient (Figure 10).

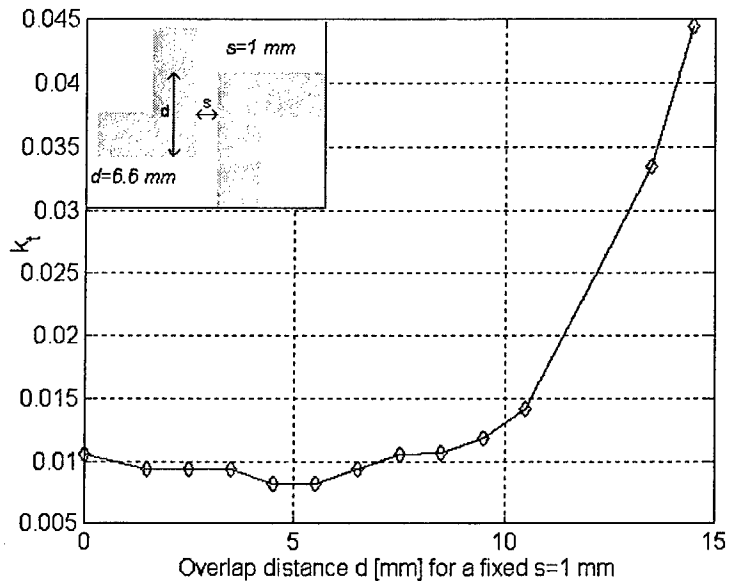


Figure 10. Coupling coefficient of the double-element antenna versus overlap distance, d , obtained from full-wave simulations

Type	Size	Bandwidth	Gain			
Double slot	$0.165\lambda_0 \times 0.157\lambda_0$	2.4%	f (MHz)	848	852	860
			G (dB)	1.5	1.7	1.7
Single-Slot	$0.133\lambda_0 \times 0.154\lambda_0$	0.9%	0.8 dB			

TABLE III. COMPARISON OF DIFFERENT PARAMETERS BETWEEN THE SINGLE ELEMENT ANTENNA AND DOUBLE-ELEMENT ANTENNA.

The radiation patterns of the antenna were measured in the anechoic chamber and the H- and E-plane co- and cross polarized radiation patterns are shown in Figure 11. It is seen that the cross-polarization level is negligible for both cuts. The two slot antennas in the double-element antenna are excited in phase (even mode) therefore their equivalent magnetic currents generate in-phase fields which will add up in the far field. Therefore the overall directivity of the double-element miniaturized antenna is expected to be more than a single element and indeed the measurements verify that (Table III). Gain measurements were performed using a standard log-periodic antenna and it was found that the double-element antenna has a gain of at least 1.5dB whereas the gain of a single element antenna is just 0.8 dB. The current distribution, however, is not the same for different frequencies therefore slight changes in radiation pattern and directivity occur over the bandwidth. The double-element slot antenna is about 25% larger than the single element antenna and features a bandwidth which is 2.7 times that of the single element antenna. In comparison with [11] which uses a folded slot antenna topology to increase the bandwidth, the increase in size is smaller (25% vs. 34%)

and the increase in bandwidth is larger (2.7 vs. 2). If the size of the double-element antenna is reduced to the size of the single antenna a bandwidth increase of 100% can be achieved without increasing the size.

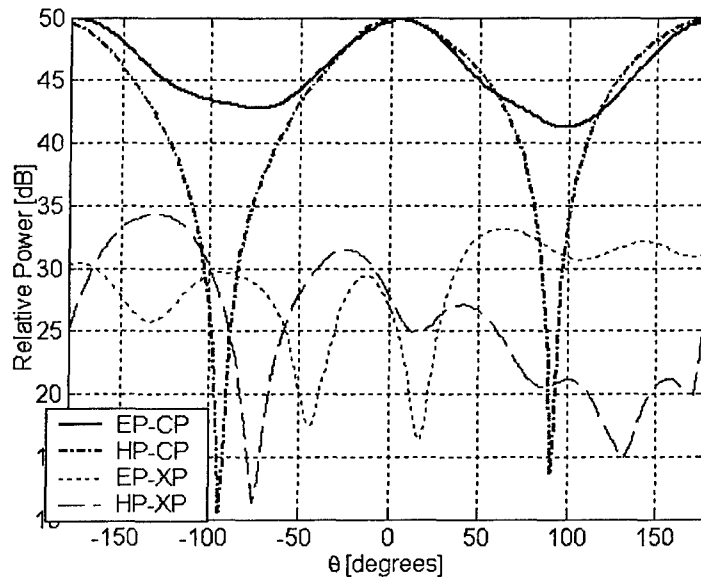


Figure 11. Radiation patterns of the double-element miniaturized slot antenna

5. CONCLUSIONS

Two different methods to enhance the bandwidth of slot antennas were introduced. It was demonstrated that a wide slot antenna, when fed with a narrow microstrip feed, shows a fictitious resonance at a frequency above the first and below the second resonance frequencies. This can be exploited to achieve a high bandwidth or a dual band antenna with similar radiation patterns for both bands.

Parasitic coupling technique when applied to a miniaturized antenna results in an antenna with a very small electrical size and considerable bandwidth compared to equivalent resonant patch antennas. Bandwidth of up to 3% can be achieved for electrical sizes as small as $0.15\lambda_0 \times 0.15\lambda_0$.

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